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Hybrid Concatenated Coding Scheme for MIMO Systems

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ABSTRACT

In this paper, two hybrid concatenated super-orthogonal space-time trellis codes (SOSTTC) applying iterative decoding are proposed for flat fading channels. The encoding operation is based on the concatenation of convolutional codes, interleaving and super-orthogonal space-time trellis codes. The first concatenated scheme consists of a serial concatenation of a parallel concatenated convolutional code with a SOSTTC while the second consists of parallel concatenation of two serially concatenated convolutional and SOSTTC codes. The decoding of these two schemes is described, their pairwise error probabilities are derived and the frame error rate (FER) performances are evaluated by computer simulation in Rayleigh fading channels. The proposed topologies are shown to perform better than existing concatenated schemes with a *constituent* code of convolutional and space-time codes in literature.

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1. INTRODUCTION

The information capacity gain of a wireless system can be increased by employing multiple transmit and/or receive antennas in a communication system [1-5]. Space-time coding lead to an increase in both bandwidth efficiency and reliability of wireless communication channels by combining spatial and temporal diversity [6]. Super-orthogonal space-time trellis code (SOSTTC) is the recently introduced space-time code that offers improved performance over earlier space-time constructions. SOSTTC combine set partitioning based on the coding gain distance and a super set of space-time block codes in a systematic way to offer full diversity and improved coding gain [7-9].

The invention of turbo coding with its astonishing performance has attracted the interest of researchers to the subject of concatenated coding schemes in recent times. Turbo codes which are built from parallel concatenation of convolutional codes with iterative decoding perform close to the Shannon limit in additive white Gaussian noise (AWGN) channels [10]. Serially concatenated convolutional codes were investigated in [8] with the turbo principles while in [9] hybrid concatenated convolutional codes with a Soft-input Soft-Output (SISO) maximum *a posteriori* decoding module was proposed.

The use of channel coding with space-time codes (STC) has the advantage of providing additional time diversity especially in fast fading channels. Various concatenated topologies have been proposed in literature with reported improved performance over conventional STC [13-21]. In [18] serial concatenation of convolutional codes and space-time trellis codes (STTC) was proposed while in [22] a double concatenated topology consisting of a parallel concatenated convolutional code (PCCC) with an inner STTC was proposed. In [23], the hybrid concatenated STC applying iterative decoding was analyzed. It was shown that choosing recursive codes as the constituent codes results in higher coding gain. Concatenation involving the convolutional code (CC) and SOSTTC was investigated over flat fading channels in [24-26].

In order to improve the performance of SOSTTC, two concatenated SOSTTC topologies over flat fading channels are proposed in this paper. The first consists of a parallel concatenated convolutional code concatenated serially with an inner SOSTTC (PC-SOSTTC) while the second, called hybrid concatenated SOSTTC (HC-SOSTTC), involves hybrid concatenated convolutional codes (HCCC) and inner SOSTTC codes. Simulations results are presented for the case of two transmit and one receive antenna in quassi static and fast fading Rayleigh channels. Both recursive and non-recursive convolutional codes were considered as the outer codes and the results are presented in terms of frame error rate (FER).

The paper is organized as follows. Firstly, the system model of the proposed scheme consisting of the channel model, the encoder and the decoder structures is described. Then the pairwise error probability of the concatenated schemes is presented. Thereafter, the performance of the concatenated scheme is evaluated by computer simulations and finally the paper is concluded.

2. RESEARCH METHOD

2.1 Parallel Concatenated-Super Orthogonal Space-Time Trellis Code

2.1a Encoder

The block diagram of the PC-SOSTTC encoder is shown in Figure 1 where the input bits are encoded by convolutional code 1 (CC1) as well as by convolutional code 2 (CC2) after interleaving by interleaver π_p. All the output bits from CC1 and CC2 are converted to a single serial stream. The serial stream is then interleaved by π_s and finally SOSTTC encoded to produce the complex symbols that are transmitted according to the SOSTTC transmission matrix at each of the transmit antennas. The interleavers are all pseudo-random and operate on bits and not symbols. The convolutional encoders are either both recursive systematic convolutional (RSC) or non-recursive convolutional (NRC) encoders. All the encoders are terminated with tail bits and to ensure uncorrelated fading and all the antennas are well separated.

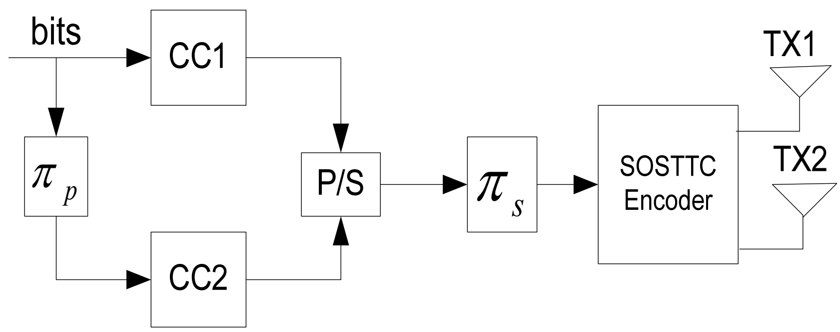

Figure 1. Encoder block diagram of the PC-SOSTTC system

2.1b Decoder

In this section, the iterative decoding process of the proposed concatenated schemes is described. The received signal after match filtering at antenna j $(j=1,\ldots,n_R)$ at time instant t is a noisy superposition of the n_T transmitted signals given by

$$r_t^j = \sum_{i=1}^{n_T} \rho_t^{i,j} s_t^i + \eta_t^j \tag{1}$$

where $\rho_t^{i,j}$ is the path gain from transmit antenna i to receive antenna j, η_t^j is the additive noise, n_T is the number of transmit antennas and s_t^i represent the Quadrature Phase Shift Keying (QPSK) symbols transmitted through the i_{th} transmit antenna, $i=1,2$ at time t. Both $\rho_t^{i,j}$ and η_t^j are modeled as independent samples of the zero mean complex Gaussian random variable with variance 0.5 and $N_0/2$ respectively per dimension. The signal to noise ratio (SNR) is defined per receive antenna as E_b/No, where E_b is the energy per bit. The flat Rayleigh fading channel is considered and the fading is statistically independent from one transmit-receive antenna pair to any other.

The PC-SOSTTC decoder (Figure 2) employs three SISO modules decoder that exchange soft information iteratively between themselves. The symbol-by-symbol maximum *a posteriori* (MAP) decoder is used for the inner SOSTTC decoder and a bit-by-bit MAP decoder is used for the outer convolutional decoder. The four ports of the SISO system of the CCs are used in the iterative decoding of the two outer decoders in order to fully exploit the potentials of the *a posteriori* probability (APP) algorithm.

Figure 2 shows a simplified diagram for the PC-SOSTTC decoder. For the purpose of simplification of description, the subscript *t* of λ and the superscript *j* of *c* and u are dropped. The decoder is specified by the subscript of *c* or *u* where the SOSTTC encoder is represented by *st,* CC1 is represented by 1 and CC2 is represented by 2. Since *a priori* information is unavailable on the first iteration, the SISO inputs $\lambda(u_{st},I)$, $\lambda(u_1,I)$ and $\lambda(u_2,I)$ are all set to zero. The coded intrinsic log likelihood ratio (LLR) for the SOSTTC SISO module is computed as shown in (2)

$$\lambda(c_{ST},I)=-\frac{1}{2\sigma^2}\sum_{j=1}^{n_R}\left|r-\sum_{i=1}^{n_T}\rho_t^{i,j}s_t^i\right|^2+\frac{1}{2\sigma^2}\sum_{j=1}^{n_R}\left|r-\sum_{i=1}^{n_T}\rho_t^{i,j}s_0^i\right|^2 \quad (2)$$

where s_0^i is the reference symbol, n_R is the number of receive antennas and σ^2 is the variance of the AWGN. The SOSTTC SISO takes the intrinsic LLR $\lambda(c_{st},I)$ and the *a priori* information from both the CC1 SISO and CC2 SISO which are initially set to zero and compute the extrinsic LLR $\hat{\lambda}(u_{st},O)$. The extrinsic LLR is then passed to the inverse interval π_s^{-1} from where the information pertaining to the coded bits of CC1 and CC2, $\lambda(c_1,I)$ and $\lambda(c_2,I)$ respectively, are extracted. The output LLRs $\lambda(c_1,O)$ and $\lambda(u_1,O)$ are calculated by the CC1-SISO. The LLR $\lambda(u_1,I)$ is subtracted from $\lambda(u_1,O)$ to obtain the LLR $\tilde{\lambda}(u_1,O)$ which is sent through the interleaver π_p to obtain the intrinsic information $\lambda(u_2,I)$ for the CC2-SISO. The output LLRs $\lambda(c_2,O)$ and $\lambda(u_2,O)$ are also calculated by the CC2-SISO. The LLR $\lambda(u_2,I)$ is subtracted from $\lambda(u_2,O)$ to obtain the LLR $\tilde{\lambda}(u_2,O)$ which is then sent via the de-interleaver π_p^{-1} to obtain the intrinsic information $\lambda(u_1,I)$ for the CC1-SISO. A single LLR stream constructed from $\tilde{\lambda}(c_2,O)$ and $\tilde{\lambda}(c_1,O)$ is interleaved by π_s to become $\lambda(u_{st},I)$ on the next iteration. The LLR $\lambda(u_2,O)$ is interleaved (π_p) to obtain $\tilde{\lambda}(u_2,O)$ which is added to $\lambda(u_1,O)$ on the final iteration upon which the decision device acts to determine the input bits.

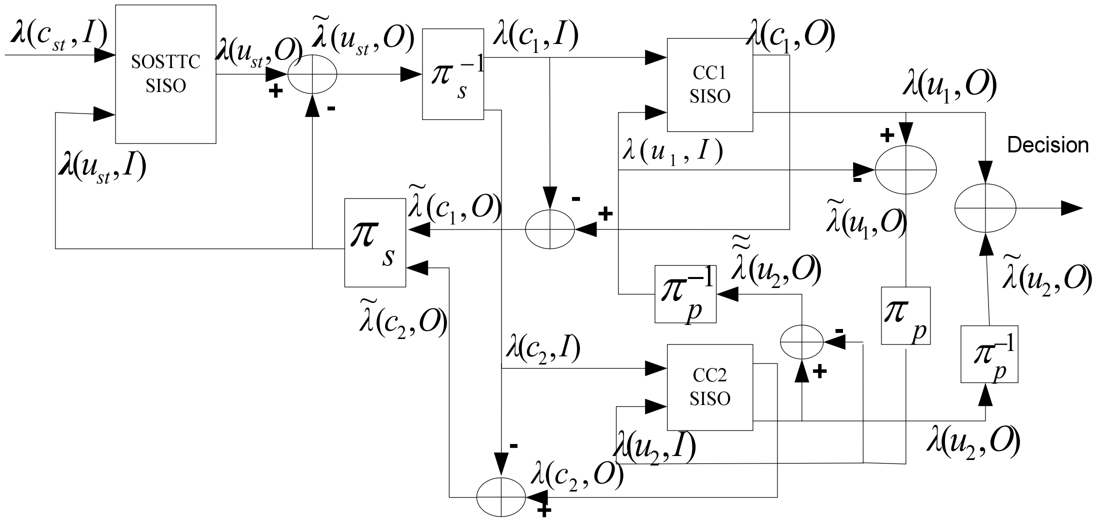

Figure 2. Decoding block diagram of the PC-SOSTTC system

2.2 Hybrid Concatenated Super-Orthogonal Space-Time Trellis Code

2.2a Encoder

In Figure 3, the transmitting block diagram of the HC-SOSTTC system is shown. The HC-SOSTTC topology consists of a parallel concatenation of two serially concatenated schemes. Each of the serial concatenated schemes consists of an outer convolutional code concatenated via an interleaver with an inner SOSTTC encoder. In the system, a block of *N* independent bits is encoded by the convolutional outer encoder, CC1, of the upper serial part of the scheme. The output of the upper convolutional encoder is then passed through a random bit interleaver (π_1). The permuted bits from the interleaver are then fed to the upper SOSTTC encoder to generate a stream of complex data that are transmitted from each of the transmit antennas using the SOSTTC transmission matrix.

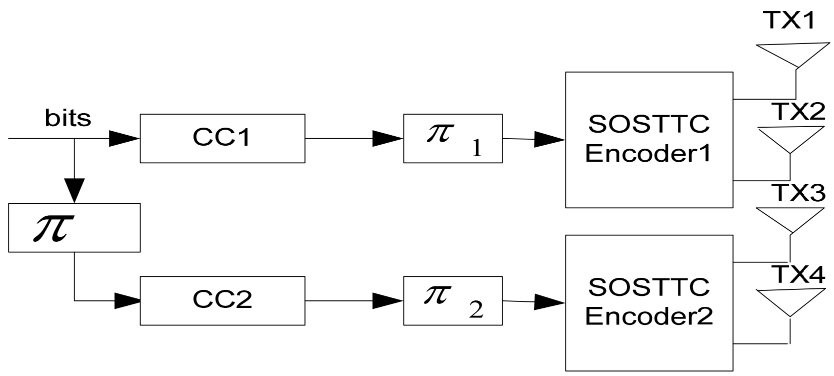

Figure 3. Encoder block diagram of the HC-SOSTTC system

In the lower serial part of the encoding, the lower convolutional encoder, CC2, receives the permuted version of the block of *N* independent bits and generates blocks of coded bits which are passed through another interleaver (π_2) to the lower SOSTTC encoder. The complex data from the output of the lower SOSTTC encoder are transmitted from the transmit antennas. It should be noted that the same convolutional and SOSTTC codes are used in the upper and lower systems. Each of the encoders is terminated using appropriate tail bits. All the four transmit antennas are well separated to ensure uncorrelated fading.

2.2b Decoder

The HC-SOSTTC decoder consists of two serial arms and one parallel sector as shown in Figure 4. The decoder is specified by the subscript of *c* or *u,* where for the upper SOSTTC encoder *st1* is used, and *st2* is used for the lower SOSTTC encoder, 1 is used for the upper convolutional encoder, CC1, while 2 is used for the lower convolutional encoder, CC2. The coded intrinsic LLR for the SOSTTC SISO module is computed as in (2).

The SOSTTC1 SISO takes the intrinsic LLR $\lambda(c_{st1}, I)$ and the *a priori* information from the CC1 SISO which is initially set to zero and compute the extrinsic LLR $\tilde{\lambda}(u_{st1}, O)$. This extrinsic LLR from the SOSTTC1 SISO is passed through the de-interleaver (π_1^{-1}) to obtain $\lambda(c_1, I)$. The LLR's output of the CC1 SISO module which are $\lambda(c_1, O)$ and $\lambda(u_1, O)$ are calculated. The LLR $\boldsymbol{\lambda}(c_1, I)$ is subtracted from $\boldsymbol{\lambda}(c_1, O)$ to obtain the LLR $\tilde{\boldsymbol{\lambda}}(c_1, O)$ which is then sent via interleaver π_1 to obtain the intrinsic information $\lambda(c_{st1}, I)$ for the SOSTTC1-SISO for the next iteration.

For the lower parallel arm, the SOSTTC2 SISO takes the intrinsic LLR $\lambda(c_{st2}, I)$ and the *a priori* information from the CC2 SISO which is also initially set to zero and compute the extrinsic LLR $\tilde{\lambda}(u_{st2}, O)$. This extrinsic LLR from the SOSTTC2 SISO is passed through the de-interleaver ($\boldsymbol{\pi}_2^{-1}$) to obtain $\lambda(c_2, I)$. The LLRs $\lambda(c_2, O)$ and $\lambda(u_2, O)$ from the output of the CC2 SISO module is then calculated. The LLR $\boldsymbol{\lambda}(c_2, I)$ is subtracted from $\boldsymbol{\lambda}(c_2, O)$ to obtain the LLR $\tilde{\boldsymbol{\lambda}}(c_2, O)$ which is then passed through the interleaver $\boldsymbol{\pi}_2$ to obtain the intrinsic information $\lambda(c_{st2}, I)$ for the SOSTTC2-SISO.

For the parallel interconnection component of the iterative decoding process, the LLR $\tilde{\lambda}(u_2, O)$ obtained by subtracting the LLR $\lambda(u_2, I)$ from the LLR $\lambda(u_2, O)$ is sent via the de-interleaver $\boldsymbol{\pi}^{-1}$ to obtain the LLR $\lambda(u_1, I)$ which is the uncoded *a priori* information passed from the CC2 SISO into the CC1 SISO. Also the LLR $\tilde{\lambda}(u_1, O)$ obtained by subtracting LLR $\lambda(u_1, I)$ from the LLR $\lambda(u_1, O)$ is sent via the interleaver $\boldsymbol{\pi}$ to obtain the LLR $\lambda(u_2, I)$ which is the uncoded *a priori* information passed from the CC1 SISO into the CC2 SISO.

The process is iterated several times and the bit with the maximum APP is chosen by the decision device in the last iteration using the summed values of the output uncoded LLRs of both the CC1and CC2 SISO decoders.

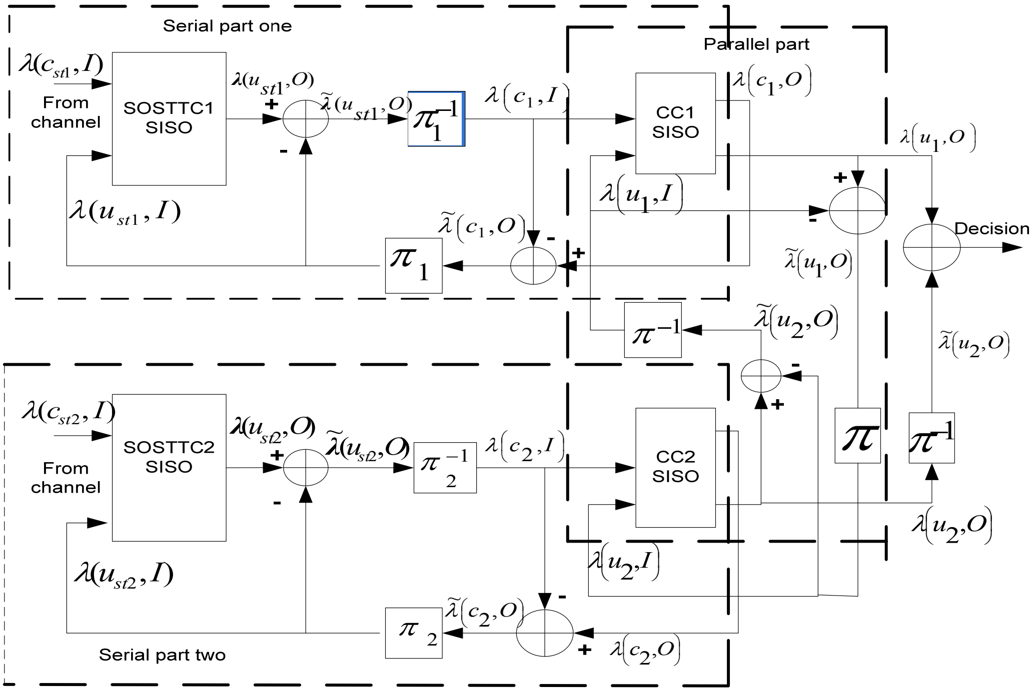

Figure 4. Decoding block diagram of the HC-SOSTTC system

3. PAIRWISE ERROR PROBABILITY (PEP) ANALYSIS

In this section, the performance bound for the concatenated schemes is derived for the case of quasi-static and fast fading channels. For slow fading, the entire frame is subjected to the same fade while for fast fading, the symbols within the frames are assumed to be subjected to independent fades.

3.1 Pairwise Error Probability for Quasi-Static Fading Channels

Let the transmitted code word and the erroneously decoded code word be denoted by $\boldsymbol{c}$ and $\hat{\mathbf{c}}$ respectively. If the symbol-wise Hamming distance between $\boldsymbol{c}$ and $\hat{\mathbf{c}}$, is denoted by $d(\mathbf{c}, \hat{\mathbf{c}})$ and assuming maximum likelihood (ML) decoding, the conditional pair-wise error probability PEP that the receiver will select $\hat{\mathbf{c}}$ over $\boldsymbol{c}$ conditioned on the channel gains and assuming perfect channel state information CSI at the receiver, is given by (3)

$$P(c \to \hat{c}) = Q\left(\sqrt{\frac{E_s d^2}{2N_0} \sum_{i=1}^{n_T} \sum_{j=1}^{n_R} \left|h_{i,j}\right|^2}\right) \tag{3}$$

where

$$d^2 = \sum_{l=1}^{d(c,\hat{c})} \left|c(l) - \hat{c}(l)\right|^2 \tag{4}$$

is the squared Euclidean distance of the outer code.

By using $Q(x) = \exp(-\frac{x^2}{2})$, we have

$$P(\mathbf{c} \to \hat{\mathbf{c}} \mid \mathbf{H}) = \exp\left(-\frac{E_s d^2}{4N_0} \sum_{i=1}^{n_T} \sum_{j=1}^{n_R} \left|h_{i,j}\right|^2\right) \tag{5}$$

Y is defined as

$$\mathrm{Y} = \sum_{i=1}^{n_T}\sum_{j=1}^{n_R}\left|h_{i,j}\right|^2 \tag{6}$$

which is a chi-squared distributed random variable, each having $2n_T n_R$ degrees of freedom with the probability distribution function (pdf) given as

$$P_{\mathrm{Y}}(y) = \frac{1}{(n_T n_R - 1)!}\, y^{(n_T n_R - 1)} e^{-y}, \qquad y > 0 \tag{7}$$

In order to compute the average PEP, we average (5) with respect to the distribution of Y ,

$$P(\mathbf{c}\to\hat{\mathbf{c}}) = \int_0^{\infty} \exp\left(-\frac{E_s d^2}{4N_0} y \frac{1}{(n_T n_R - 1)!} y^{(n_T n_R - 1)} e^{-y}\right) dy \tag{8}$$

Using the integral function [23]

$$\int_0^{\infty} x^n e^{-\mu x} dx = n!\,\mu^{-n-1} \tag{9}$$

we have

$$P(\mathbf{c}\to\hat{\mathbf{c}}) = \left(1 + \frac{E_s d^2}{4N_0}\right)^{-n_T n_R} \tag{10}$$

At high SNR, (10) can be approximated as

$$P(\mathbf{c}\to\hat{\mathbf{c}}) \approx \left(\frac{E_s d^2}{4N_0}\right)^{-n_T n_R} = \left(d^2(c,\hat{c})\frac{E_s}{4N_0}\right)^{-n_T n_R} \tag{11}$$

Equation (11) indicates that the diversity order of $n_T n_R$ is achieved in a quasi-static fading channel. For the PC-SOSTTC system, therefore, the diversity order of 2 is achievable while for the HC-SOTTC, the diversity order of 4 is achievable.

3.2 Pairwise Error Probability for Fast Fading Channels

In the case of a fast fading channel, the conditional PEP that the receiver will select code word $\hat{\mathbf{c}}$ over $\boldsymbol{c}$ assuming that CSI is known at the receiver and conditioned on the channel gain, is given by [6]

$$P(\mathbf{c}\to\hat{\mathbf{c}}\,|\,\mathbf{H}) = Q\left(\sqrt{\frac{E_s}{2N_0}\sum_{k=1}^{d(c,\hat{c})}\sum_{i=1}^{n_T}\sum_{j=1}^{n_R}\left|hi,j\right|^2\left|c(k)-\hat{c}(k)\right|^2}\right) \tag{12}$$

where $\left|c(k)-\hat{c}(k)\right|^2$ is the normalized squared Euclidean distance between the correct path signal and the error path signal at time index *k*. By using $Q(x) = \exp(-\frac{x^2}{2})$, we have

$$P(\mathbf{c} \rightarrow \hat{\mathbf{c}} \mid \mathbf{H}) = \prod_{k=1}^{d(c,\hat{c})} \left\{ \exp\left(-d_k^2 \sum_{j=1}^{n_R} \sum_{i=1}^{n_T} \left| h_{i,j}(k) \right|^2 \right) \right\} \tag{13}$$

where

$$d_k^2 = \frac{E_s}{4N_0} \left| c(k) - c(k) \right|^2 \tag{14}$$

Y_k is defined as

$$\mathrm{Y}_k = \sum_{j=1}^{n_R} \sum_{i=1}^{n_T} \left| h_{i,j}(k) \right|^2 , for\, k = 1, 2, ..., d(c,\hat{c}) \tag{15}$$

which are independent and chi-squared distributed, each with $2n_T n_R$ degrees of freedom with a pdf given by (7). In order to compute the average PEP, (13) is averaged with respect to the distribution of Y_k

$$P(\mathbf{c} \rightarrow \hat{\mathbf{c}}) = \prod_{k=1}^{d(c,\hat{c})} \int_0^{\infty} \exp\left(-d_k^2 y_k \frac{1}{(n_T n_R - 1)!} y_k^{(n_T n_R - 1)} e^{-y} \right) dy_k \tag{16}$$

Using (9), we have

$$P(\mathbf{c} \rightarrow \hat{\mathbf{c}}) = \prod_{k=1}^{d(c,\hat{c})} (1 + d_k^2)^{-n_T n_R} \tag{17}$$

At high SNR, (17) can be approximated as

$$P(\mathbf{c} \rightarrow \hat{\mathbf{c}}) \approx \prod_{k=1}^{d(c,\hat{c})} (d_k^2)^{-n_T n_R} \tag{18}$$

$$P(\mathbf{c} \rightarrow \hat{\mathbf{c}}) = \prod_{k=1}^{d(c,\hat{c})} (\left| c(k) - \hat{c}(k) \right|^2)^{-n_T n_R} (\frac{E_s}{4N_0})^{-n_T n_R} \tag{19}$$

From (19), it is clear that the diversity order of $n_T n_R d_{\min}$ is achieved in a fast fading channel, where d_{min} is the minimum Hamming distance of the outer convolutional code.

4. RESULTS AND ANALYSIS

In this section, simulation results are presented to evaluate the performance of the proposed concatenated scheme over Rayleigh fading channels. The performance of the two proposed topologies is evaluated over both slow and fast fading channels. Narrow band transmission is assumed. Therefore, the results illustrate the performance in time division multiple access (TDMA) type systems, like the global system for mobile communication (GSM), IS 136, or enhanced data rates for GSM Evolution (EDGE). The results are presented in terms of FER versus E_b/No. In all the simulations, 130 symbols per frame are transmitted from each of the transmit antennas. Unless otherwise stated the number of iterations of the simulation is set to six because the performance of the systems is observed to reach saturation at the sixth iteration.

The four-state, QPSK SOSTTC in [7] is considered as the inner code. For the outer code, RSC and NRC rate 1/2, four-state convolutional code are employed for the HC-SOSTTC while for the PC-SOSTTC the rate 2/3, four-state RSC and NRC convolutional codes are used. In the two architectures, the outer convolutional codes are both either RSC or NRC.

Figures 5 and 6 show the FER performance of the HC-SOSTTC and the PC-SOSTTC systems respectively, for various numbers of decoding iterations in quasi-static fading channels. The performance of both schemes is observed to improve with an increase in the number of iterations and start saturating at about the 4th iteration. As can be observed from the FER performance curve, the HC-SOSTTC achieves full diversity order of four which is consistent with the observation from the PEP analysis. The concatenation adds no additional diversity to the scheme but achieves significant coding gain as seen by the horizontal shift of the FER performance curve. Also from the FER performance curve of the PC-SOSTTC, no added diversity is achieved by the system but there is significant coding gain improvement by the concatenation.

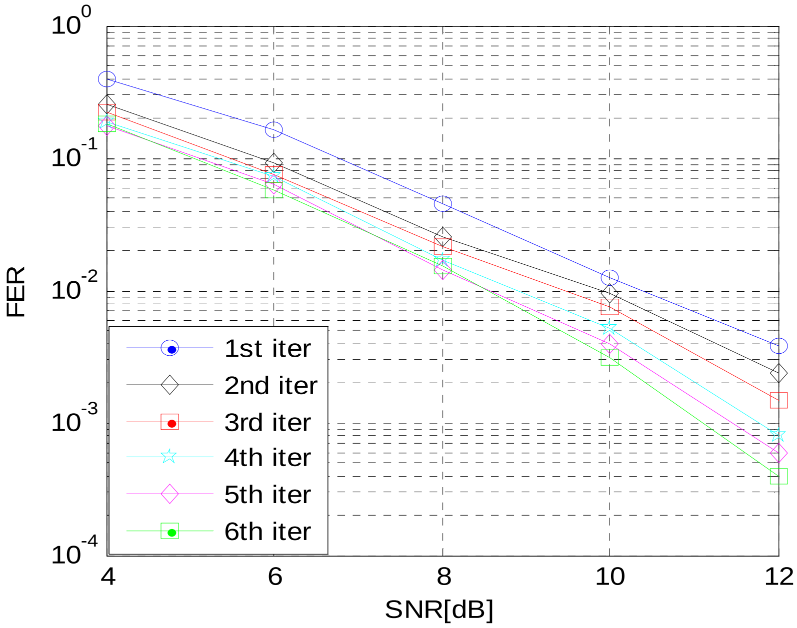

Figure 5. Frame error rate (FER) performance of HC-SOSTTC for various numbers of decoding iterations

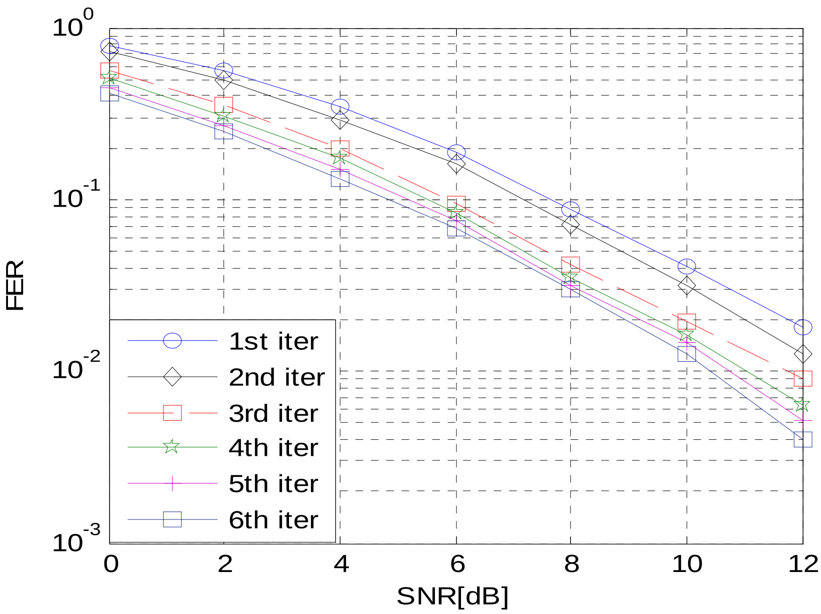

Figure 6. Frame error rate (FER) performance PC-SOSTTC for various numbers of decoding iterations

In Figure 7 the FER performance for the HC-SOSTTC comparing the case of RSC outer code, NRC outer code, RSC with STTC inner code and NRC with STTC inner code is shown. On the same plot the FER performance of super-orthogonal space-time-convolutional code (SOST-CC) scheme from Pillai and Mneney [21] is plotted for comparison. The SOST-CC concatenates convolutional code serially with SOSTTC which is equivalent to one serial arm of the HC-SOSTTC system. For the STTC inner code, the four states STTC from [6] is employed. As can be seen from the FER performance curve, the HC-SOSTTC system with RSC outer code outperformed the scheme with NRS outer code and the HC-STTC codes. The code with RSC outer code presented better coding gain when compared with the scheme with NRC outer code because recursive codes, unlike their non-recursive counterparts, achieve interleaving gain in iterative decoding. In comparison with the SOST-CC, the HC-SOSTTC outperformed it in terms of both diversity order and coding gain. The HC-SOSTTC scheme achieves a higher diversity order because of the number of transmitting antennas involved. The diversity order of SOST-CC is two while that of HC-SOSTTC is four. In terms of coding gain, the HC-SOSTTC outperformed the SOST-CC by 3.5 dB at the FER of 10^{-2}.

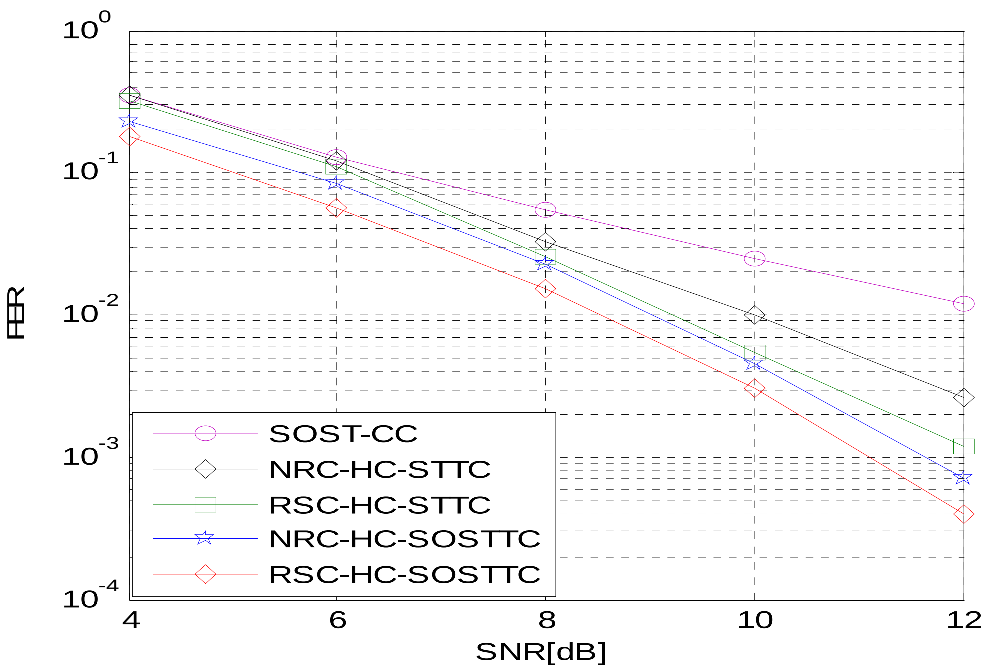

Figure 7. Frame error rate (FER) performance of HC-SOSTTC over quasi-static fading channel

Figure 8 shows a FER performance comparison for the PC-SOSTTC system with RSC outer code, NRC outer code and the PC-STTC code. As can be observed from the performance figure, the scheme with outer RSC code achieves improved coding gain over the scheme with outer NRC code. It is also observed that PC-SOSTTC achieved higher coding gains than the PC-STTC code.

The FER performance of the HC-SOSTTC over fast fading channel is shown in Figure 9. The performance of the code is evaluated over this channel condition using both NRC and RSC outer codes. The scheme's performance is also compared with the CC-SOSTTC scheme from Altunbas [24] over the same channel condition. The scheme with outer RSC code is observed from the FER plots to achieve higher coding gain when compared to the scheme with NRC outer code. In comparison with the CC-SOSTTC code, the HC-SOSTTC with outer RSC code has a coding gain advantage of about 4 dB over the CC-SOSTTC at the FER of 10^{-3}.

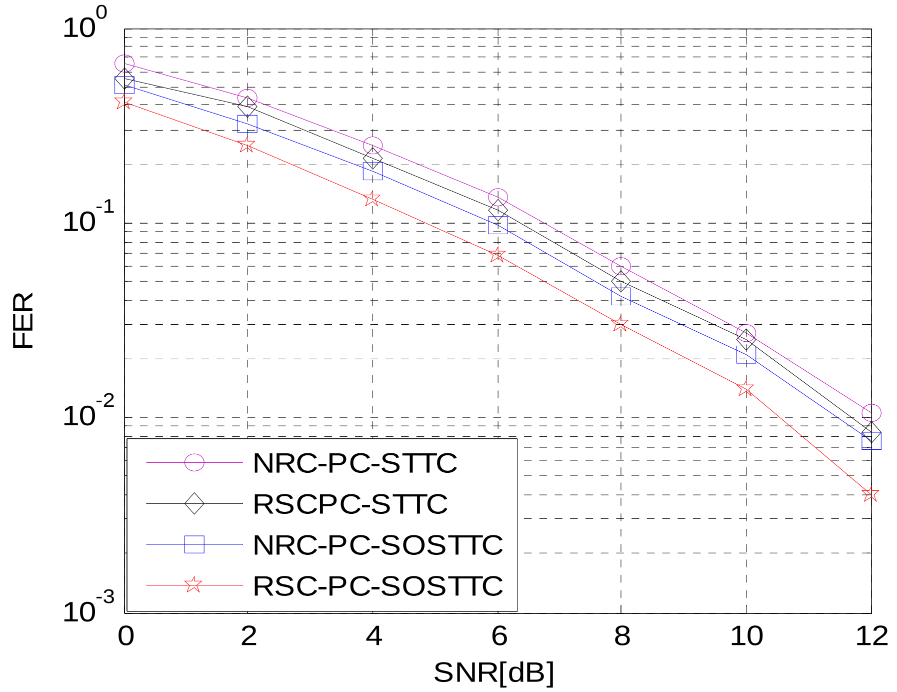

Figure 8. Frame error rate (FER) performance for PC-SOSTTC over quasi-static fading channel

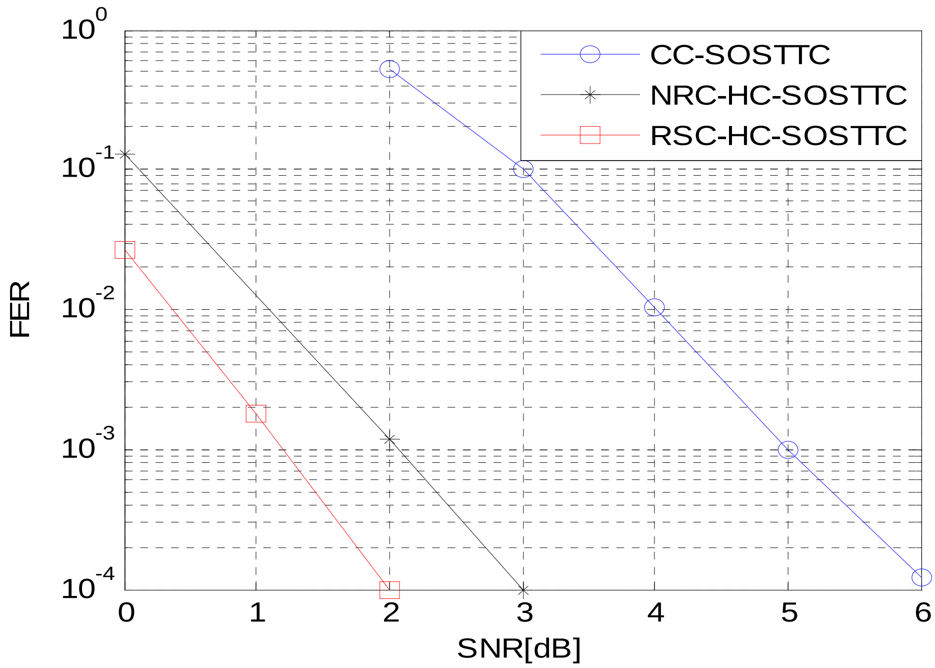

Figure 9. Frame error rate (FER) performance for HC-SOSTTC over fast fading channel

The FER performance of the PC-SOSTTC system over a fast fading channel is shown in Figure 10 using RSC and NRC outer convolutional codes. In the same figure the system with inner STTC code is shown. The PC-SOSTTC is observed to achieve a very high diversity order in a fast fading channel but experiences the error floor phenomenon. Error floor at higher SNR region is a peculiar characteristic of a parallel concatenation scheme [12]. The error floor phenomenon is more pronounced with the scheme with NRC outer code as can be observed from the FER performance plot.

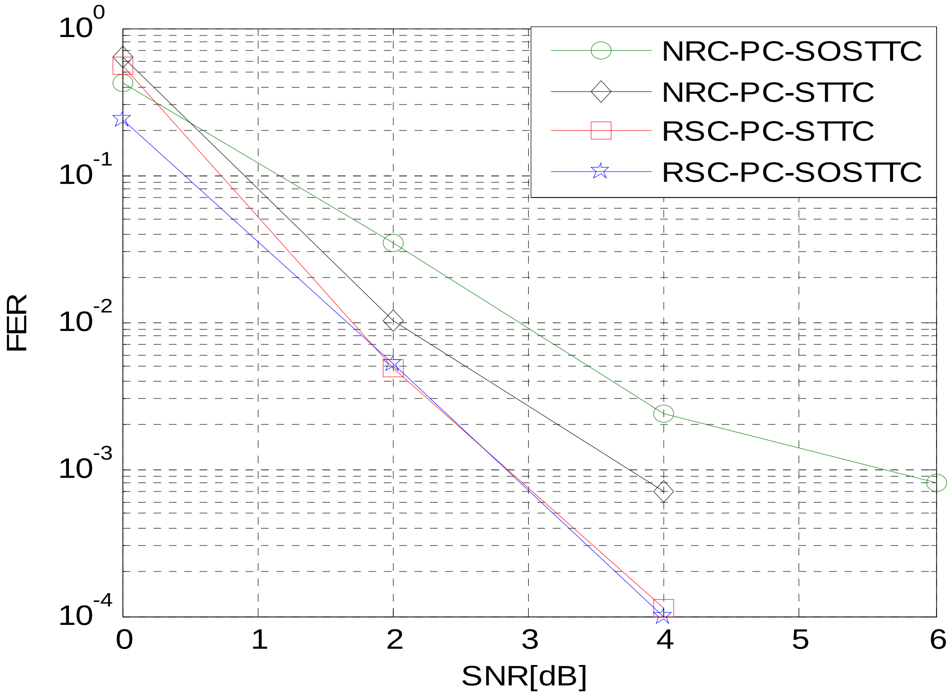

Figure 10. Frame error rate (FER) performance for PC-SOSTTC over fast fading channel

5. CONCLUSION

In this paper two concatenated schemes with constituent codes of SOSTTC and convolutional codes were proposed for MIMO systems. The encoding and the decoding of the two concatenated schemes were described and the PEP derived for the scheme for Rayleigh fading channels. Simulation results show that no additional diversity is achieved by concatenating outer code with SOSTTC in quasi-static fading channels, which is consistent with the PEP analysis. The concatenated scheme also achieves better error performance over a fast fading channel by benefiting from interleaving gain. From the literature, the HC-SOSTTC is seen to present significantly improved performance over other concatenated schemes operating over flat fading channels, but with increased complexity.

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